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DESIGN TECHNIQUES FOR VARACTOR FREQUENCY MULTIPLIERS

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by

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SUMMARY

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This paper carries through the practical design of varactor frequency doubler and tripler circuits. The basic requirements for achieving a practical frequency multiplier (an input circuit subject to power at the input frequency, an output circuit for delivering the power that has been multiplied in frequency, and the common circuitry which contains the varactor) are the basis for design. A specific design example is given in which other considerations also influence the design criteria; for example, spacecraft requirements regarding size, weight, reliability, and intended end use. The intended purpose, for the multipliers described, is to provide a local oscillator signal that is coherent both with a second local oscillator frequency and with a spacecraft transmitted frequency. The specified end use as a local oscillator requires that the output frequency spectrum be relatively free of spurious frequencies.

The local oscillator as designed consists of a cascaded tripler-doubler to provide a frequency multiplication of six.

AUTHOR

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Introduction:

This paper is concerned with the practical design of varactor frequency doublers and triplers, using manufacturers' published data, and design curves contained in Varactor Applications, authored by Penfield and Rafuse. The purpose of this paper is to illustrate a practical design technique; and, no theoretical evaluation will be attempted of either the varactor or the applicable circuits. Two complete multipliers, a single ended doubler and tripler, will be designed. The basis behind the design philosophy is the development of compact reliable circuits which will be simple (conservative in the numbers of components), stable, and easily tuned. Such circuits may not exhibit as high an efficiency or isolation of fundamental feed-through as is possible to achieve; however, they will be suitable for spacecraft equipments, can be potted, are less subject to drift and more applicable to unattended operation for extended periods.

General Multiplier Design

In general we desire to keep the reactances fairly low to achieve relatively wide bandwidths and more reliable "starting"; and, while self bias is permissible under certain conditions, we will use fixed bias which also improves starting characteristics. In spacecraft frequency multiplying systems which may be subject to under-voltage turn offs, resulting in intermittent operation or reduced drive power, a self bias system may not "start" or, if started, may exhibit very poor conversion efficiency if operating at a reduced input power. For example, circuit resonance, and a maximum use of input power, is dependent upon the average capacitance of the varactor; the proper capacitance can only be obtained with full drive power, and therefore, the circuit will contain a "residue" of series reactance which will limit input current flow. In actual practice when the reactances are relatively large, non-starting can be quite common, while with low reactances it is generally less of a problem. A combination of fixed and self bias could be considered for some applications.

In the specific design examples, we shall make considerable use of the design charts found in the volume Varactor Applications; the book explains the construction of these charts. Basically, they permit the correlation of varactor efficiency, relative drive power levels, load (or conversion) resistance presented by the varactor to frequency ω_o and, for our purposes, the equivalent internal resistance of the pseudo generator $2\omega_o$ or $3\omega_o$, to the varactor characteristics as specified by the manufacturer or as measured. The varactor parameters usually given are cut-off frequency, series resistance and maximum voltage rating (voltage breakdown). The cut-off frequency is that frequency (f_c) at which the maximum capacitive reactance (at V_{max}) equals the series resistance (r_s). The series resistance of the varactor is that resistance that would account for the power loss in a varactor that is being driven from a high frequency source, while biased in the reversed direction. The maximum voltage rating (voltage breakdown) is the maximum reverse voltage that can be applied to the diode. One other characteristic is necessary which is designated P_{norm} (normalized power), which is defined as voltage breakdown–squared divided by the varactor series resistance.

cut-off frequency (f_c)

$$f_c = \frac{1}{2\pi R_s C_{min}}$$

where: C_{min} is the varactor capacitance at voltage breakdown (V_B).

R_s = varactor series resistance

$$P_{norm} = \frac{\overline{V}_B^2}{R_s}$$

where: V_B = varactor breakdown voltage (or voltage rating).

We have the option of using the varactor in either a series or shunt configuration and of using a single ended or balanced circuit. Since the shunt configuration permits one end of the varactor to be firmly fixed, even soldered, to ground, it is to be preferred for most circuits when significant power dissipation is a consideration. This is especially helpful when using varactors with a heavy metal shoulder, such as the P.S.I. pigtail type. Single ended circuits will be used for both the doubler and tripler.

As a further consideration, an attempt will be made to match circuit input and output impedances to 50 ohms by selection of the varactor diode. In any event, if reflected power is not a problem, it may not be worthwhile to use a matching network if the impedance mismatch is less than two-to-one. The gain in efficiency may be non-existent or not worth the extra components or may adversely affect the circuit bandwidth. For such relatively small impedance mismatches it is usually simpler to alter the value of one of the resonant reactances in the basic circuit to reduce the mismatch.

In general, our purpose is to achieve a maximum ω_o current via the input circuit through the varactor; and, to achieve a maximum $2\omega_o$ or $3\omega_o$ current, as the case may be, through the varactor and the output circuit.

Design of a Single Ended Frequency Doubler

We will now design a frequency doubler to satisfy the following requirements:

Power in	–	8×10^{-3} watts
R_{in} (R_1)	–	50 ohms
R_{out} (R_2)	–	50 ohms
Input frequency (ω_o)	–	68×10^6 cycles
Output frequency ($2\omega_o$)	–	136×10^6 cycles

We also desire a small compact unit with small coils that can be slug tuned with the components mounted on a printed circuit board. The board will ultimately be potted and is required to withstand severe environmental testing. To achieve a good efficiency and keep losses low, we should attempt to use a high-f_c varactor and coils with high Q and relatively low values of reactance. The reactance values will in this case depend upon the specific varactor selected. We desire a varactor that will permit us to approach a 50-ohm circuit impedance and also to be driven fully with an input power of eight milliwatts.

Assuming a circuit resonant condition, we can find the relationship between input and load resistances with the normalized input frequency (ω_o/ω_c) and the effective varactor series resistance (R_s) as per chart

No. 1. In chart No. 1, the input and output resistances are expressed as a factor times R_s. Also, these resistances are shown both for a maximum power condition and a maximum efficiency condition.

The characteristics necessary for a full drive condition can be determined from chart No. 2. Chart No. 2 relates drive power to normalized power (P_{norm}) as a function of normalized input frequency (ω_o/ω_c). In general, the lower the operating frequency (ω_o) for a given varactor the lower the permissible drive power. Chart No. 2 also has curves of the relative power dissipated and the relative power output from a varactor doubling under full drive conditions.

For purposes of selecting a varactor, we can assume that the varactor cut-off frequency (ω_o) of available units will most likely fall between 10 gc (10^{10} cycles) and 70 gc (7×10^{10} cycles). Hence, the normalized ω_o (68×10^6 cycles) should fall between $7 \times 10^{-3}\ \omega_c$ and $10^{-3}\ \omega_c$ ($68 \times 10^6/\omega_c$). The P_{norm} required can be determined by referring to chart No. 2. From the chart we see that the input power for the estimated ω_o's, $7 \times 10^{-3}\ \omega_c$ to $10^{-3}\ \omega_c$, will range between 3×10^{-4} and 4×10^{-5} of P_{norm}.

$$P_{in} \cong 10^{-2} \text{ watts} = (3 \times 10^{-4} \text{ to } 4 \times 10^{-5})\, P_{norm}$$

$$P_{norm} = \frac{10^{-2}}{(3 \times 10^{-4} \text{ to } 4 \times 10^{-5})}$$

$$\underline{P_{norm} = 33 \text{ to } 250 \text{ watts}}$$

(as a function of normalized ω_o)

For a high ω_c (50-70 gc), P_{norm} will lie in the range of 200 to 250 watts; and, for a low ω_c (10-20 gc), P_{norm} will be in the range of 30 to 50 watts.

Note: A high ω_c will result in a low normalized ω_o.

The input and output resistances, R_1 and R_2 respectively, are a function of the R_s of the varactor and the normalized ω_o. By referring to chart No. 1 we can determine the approximate value of R_s we should have to give us our desired 50-ohm impedance. Again, ω_o is between 7×10^{-3} ω_c and 10^{-3} ω_c.

$$R_{in} = R_{out} = 50 = (15 \text{ to } 100)\ R_s$$

$$R_s = \frac{50}{15 \text{ to } 100}$$

$$\underline{R_s = 3.3 \text{ ohms to } 0.5 \text{ ohms}}$$

(for a low to high ω_c varactor respectively)

We now have sufficient information to select a varactor for our doubler circuit. Bear in mind that these values are the desired ones and they will provide a varactor that will result in the highest efficiency; however, in practice, some deviation can be permitted without too much degradation of performance. Also, the step for determining the optimum R_s can be modified if we have no compulsion regarding the avoidance of impedance matching circuits. The following table summarizes the desired parameters to aid in the selection of a varactor.

Table 1

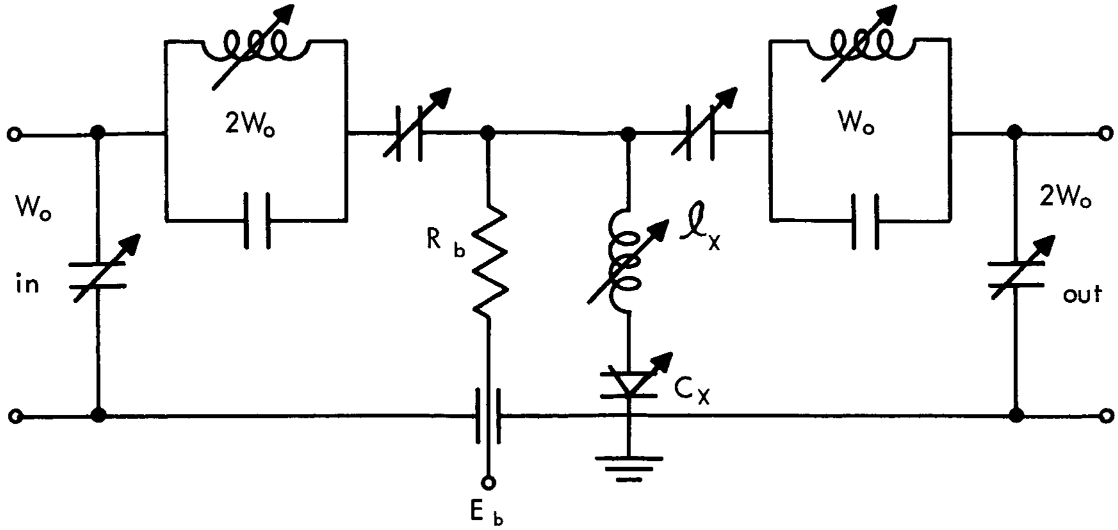
Cut-off freq. ω_c	70 gc	30 gc	10 gc
Normalized ω_o	$10^{-3}\,\omega_c$	$2.5 \times 10^{-3}\,\omega_c$	$7 \times 10^{-3}\,\omega_c$
P_{norm}	250 watts	125 watts	30 watts
R_s	0.5 ohms	1.0 ohms	3.3 ohms

Note: Again, when using the charts, a high ω_c will result in a low normalized input frequency.

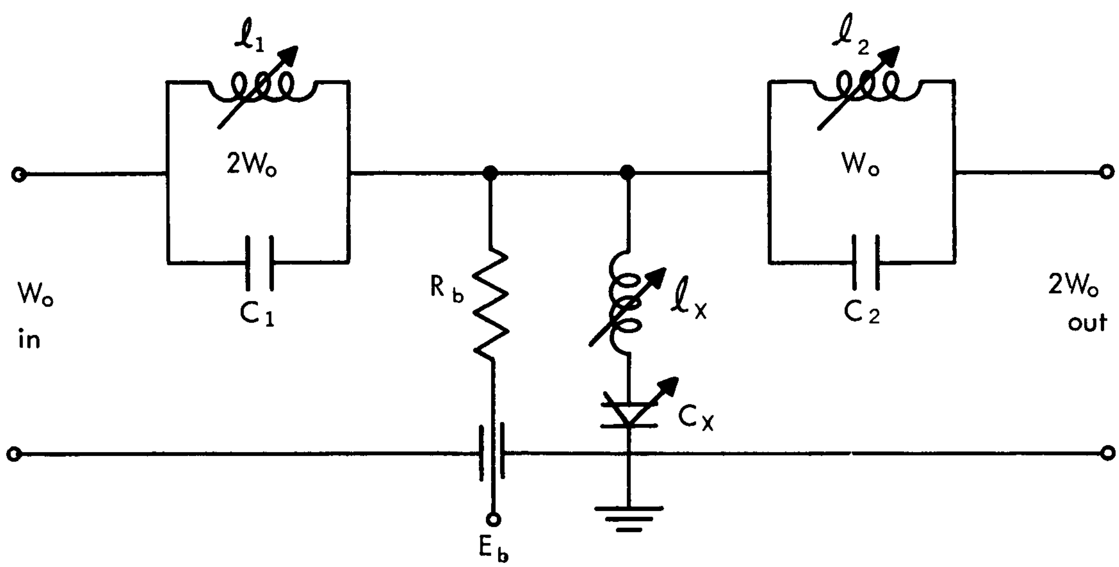
A further factor which may influence the selection of a varactor concerns the value of capacitance the varactor presents to the doubler circuitry. Circuit component values will depend on the average capacitance, C_x, of the varactor.

Figure 1 shows a basic single ended doubler configuration with a circuit simplification that we shall use. In Figure 1-b, if we make the ℓ_x-C_x leg series resonant at $\sqrt{2}\,\omega_0$ or, in our example, $\sqrt{2} \times 68 \times 10^6 = 96 \times 10^6$ cycles, then, if the inductance, ℓ_1, of the input trap (parallel resonant at $2\omega_o$) is made equal to $3/4\ \ell_x$ and the inductance, ℓ_2, of the output trap (parallel resonant at ω_0) is made equal to $3/2\ \ell_x$, the doubler circuit will be "tuned" when the proper bias is applied to the varactor. In this case the input trap ($2\omega_o$) and the $\ell_x - c_x$ combination will be series resonant to ω_o; while the output trap (ω_o) and the $\ell_x - c_x$ combination will be series resonant to $2\omega_o$.

Note: The value of C_x we will use is the average value of the varactor capacitance; its back biased value in our case. The capacitance of the abrupt junction varactor varies inversely as the square root of the applied bias voltage.



A Frequency Doubler



B. Simplified Frequency Doubler
(Varactor selected to achieve impedance match)
$\ell_1 = 3/4\,\ell_X$; $\ell_2 = 3/2\,\ell_X$
$C_1 = 2/3\,C_X$ $C_2 = 4/3\,C_X$

Figure 1–Single Ended Frequency Doubler

A doubler circuit using series resonant legs on the input and output could be used which would eliminate the need for the inductance, ℓ_x, which is in series with the varactor, c_x. However, since the intended use (local oscillator) of this doubler requires that the spectrum be relatively clean of spurious outputs it is necessary to use an output trap (ω_o), hence the requirement for ℓ_x.

The determinations of ℓ_x, ℓ_1, ℓ_2, c_1 and c_2 can be found in the appendix.

A varactor that had similar characteristics to those desired was selected from stock. It's characteristics were:

ω_c – 46 gc

ω_o – 68×10^6 cycles $\cong 1.5 \times 10^{-3}\ \omega_c$

R_s – 0.78 ohms

V_B – 18 volts

P_{norm} – $(\overline{V}_B^2/R_s) = (\overline{18}^2/0.78) = 415$ watts

C_{18v} – 4.4 $\mu\mu f$

$C_{operating}$ – $C_{18}\sqrt{(V_B/V_{op})} = 4.4\sqrt{(18/6)} \cong 7.6\ \mu\mu f$
(operating bias $\cong 18/3 \cong 6v$; Chart #4)

Referring to Chart No. 1 and Chart No. 2, the following circuit conditions can be determined:

1. optimum P_{in} (full drive)

   (Chart #2 – $\omega_o = 1.5 \times 10^{-3}\ \omega_c$)

   $P_{in} \cong 4 \times 10^{-5}\ P_n$

   $= 4 \times 10^{-5} \times 415$

$$P_{in} \cong 16 \times 10^{-3} \text{ watts}$$

(actual $P_{in} \cong 10 \times 10^{-3}$ watts)

The varactor efficiency with this drive power will be reduced perhaps 1-1/2 to 2 percent from that for full drive power as seen from Chart No. 3. Assuming the input and output impedances are matched, the varactor efficiency for this reduced drive would be approximately 95%.

2. input and output impedances

(Chart #1 – $\omega_o = 1.5 \times 10^{-3}\ \omega_c$)

$$R_{in} = 55 \cdot R_s = 55 \cdot 0.78 = 43 \text{ ohms}$$

$$R_{out} = 94 \cdot R_s = 94 \cdot 0.78 = 73 \text{ ohms}$$

$$(\max P_o)\ R_{in} = R_{out} = 70 R_s = 70 \cdot 0.78 \cong 56 \text{ ohms}$$

These values of R_{in} and R_{out} will result in minimum varactor losses. However, a 50-ohm drive and load impedance will affect varactor efficiency only slightly; and, since circuit losses will exceed the varactor losses by several times, the effect of the impedance on varactor efficiency at this frequency ($1.5 \times 10^{-3}\ \omega_c$) is minor. The circuit losses due to the expected impedance mismatch can be estimated to be about 2-3%.

Now we shall determine the circuit component values and estimate circuit losses. In general, we should attempt to keep reactances low; however, we shall be bound in the choice of circuit values by the average value of the varactor capacitance.

Note: Regarding the choice of average varactor capacitance (bias value), the reactance at ω_o should be kept within a range of perhaps 150 to 800 ohms for the best choice of circuit component values. At either very high frequencies or very low frequencies relative to ω_c it may not be possible to remain within this range.

Assuming the use of good quality capacitors, the inductors will be the source of the greater portion of the circuit losses. Referring to circuit B of Figure 1, ℓ_x resonates with C_x (varactor) at $\sqrt{2}\,\omega_o$ (96 mcs). C_x will be reversed biased at 0.35 V_B (approximately 6 volts).

Note: If a full voltage swing (full drive power) is not applied to the varactor, a lesser or greater bias can be used to effect a change in the average value of capacitance. Also, if an inadequate bias is applied, the varactor will conduct on the positive peaks of the input signal and will attempt to bias or clamp itself accordingly. Under this partial self bias condition, varying drive levels will cause the bias to vary, changing the average value of capacitance and detuning the input and output circuits. This could possibly be a useful feature; for example, if a degree of self limiting is desired.

Since the average value of C_x was determined to be 7.6 $\mu\mu$f;

$$\sqrt{2}\,\omega_o = \frac{1}{\sqrt{\ell_x C_x}} = 2\pi \cdot 96 \times 10^6$$

$$\underline{\ell_x = 0.36\ \mu h}$$

$$\ell_1 = 3/4\,\ell_x \qquad \underline{\ell_1 = 0.27\ \mu h}$$

$$\ell_2 = 3/2\,\ell_x \qquad \underline{\ell_2 = 0.54\ \mu h}$$

Now with the values of the inductances known, the traps are resonated by the proper values of C_1 and C_2

$$C_1 \cong 5.0\ \mu\mu f \quad ; \quad C_2 \cong 10.0\ \mu\mu f$$

Checking input and output reactances:

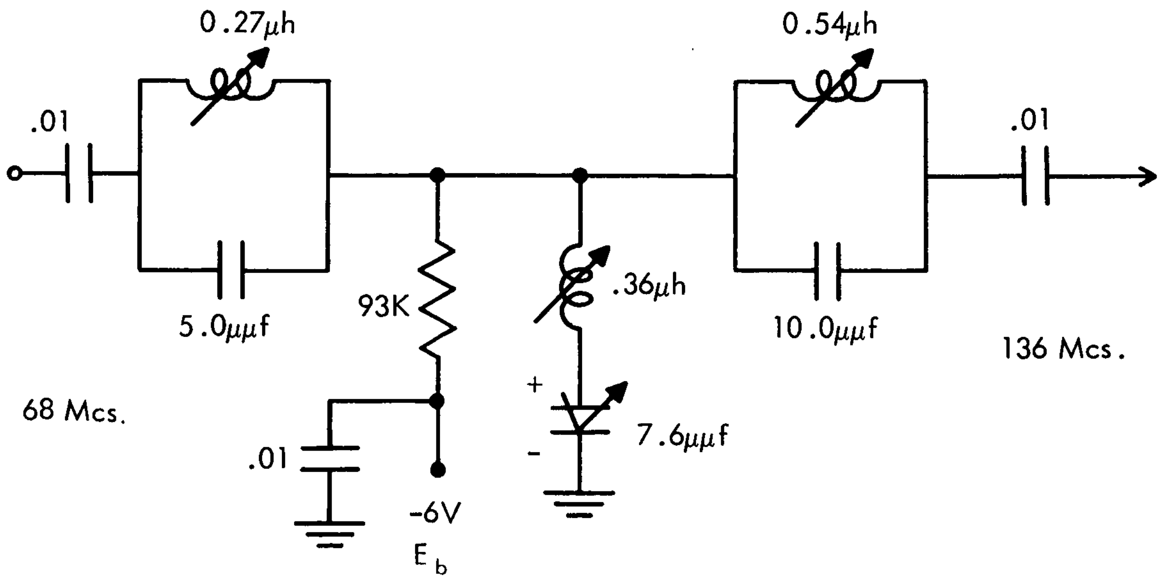


Figure 2–Component Values–Doubler

Input reactance (ω_o)

$0.27\ \mu h\ = +j\ 115$

$5.0\ \mu\mu f\ = -j\ 470$

$0.36\ \mu h\ = +j\ 152$

$7.6\ \mu\mu f\ = -j\ 305$

$$\frac{(+j\ 115)\ (-j\ 470)}{+j\ 115 - j\ 470} + j\ 152 - j\ 305 = \text{``0''}$$

$$+j\,152 + j\,152 - j\,305 \cong \text{“0”}$$

The input loop is in a series resonant condition.

Output reactance ($2\omega_o$)

$7.6\ \mu\mu f = -j\,155$

$0.36\,\mu h = +j\,305$

$10.7\ \mu\mu f = -j\,110$

$0.51\,\mu h = +j\,435$

$$-j\,155 + j\,305 + \frac{(-j\,110)\,(+j\,435)}{-j\,110 + j\,435} \cong \text{“0”}$$

$$-j\,155 + j\,305 - j\,147 \cong \text{“0”}$$

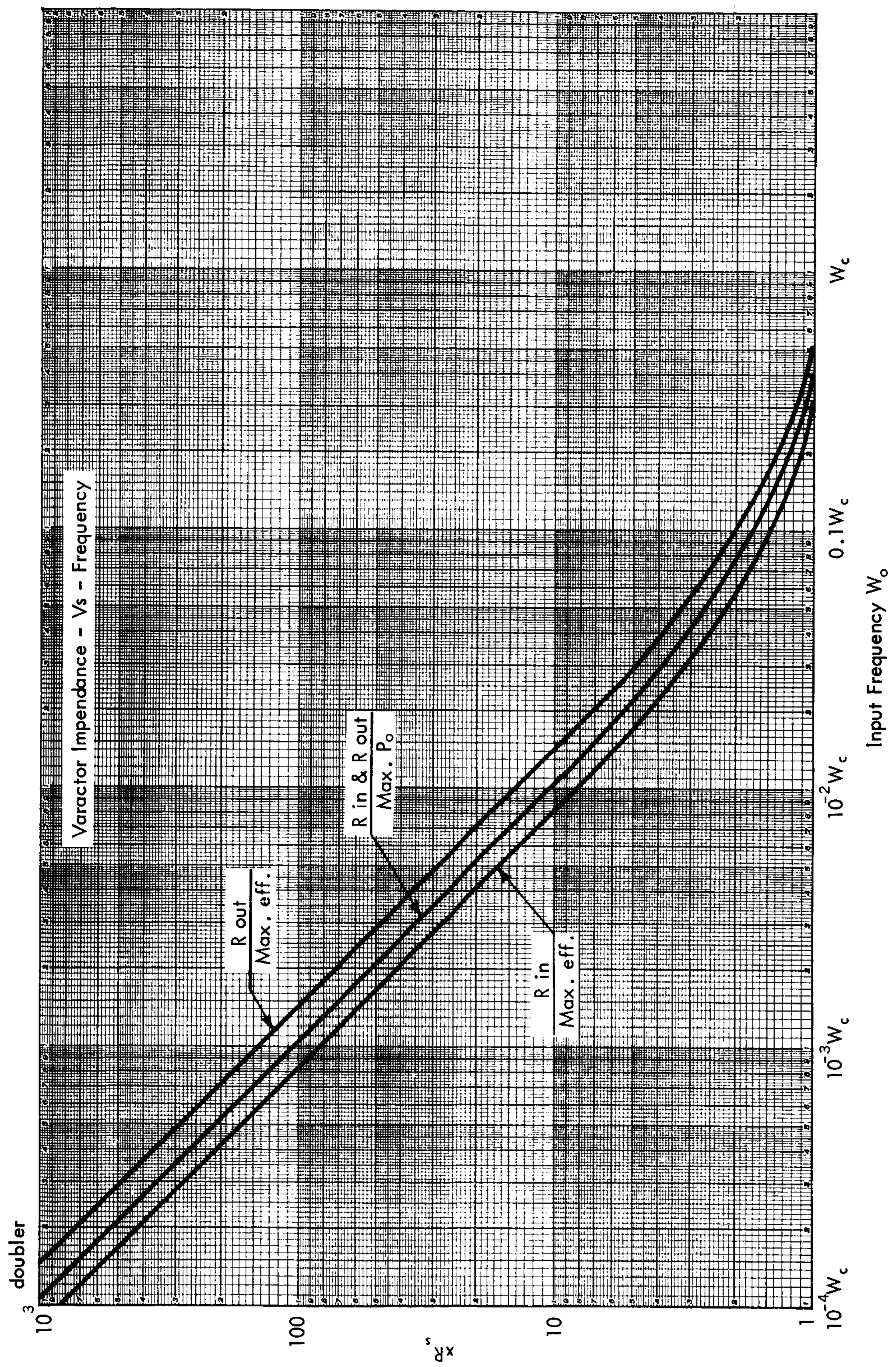
The output loop is in a series resonant condition.

For actual doubler construction we shall use small (1/4-inch diameter) coils that are slug tuned. To estimate circuit losses we assume coil Q's of 100 and a circuit impedance of 50 ohms. The various coil reactances are given above. The input loss should be approximately 5% for the input loop plus 2-3% for the ω_o trap, totaling approximately 8%. The output ($2\omega_o$) loss will be approximately 15% for the output loop plus approximately 4% for the $2\omega_o$ trap, totaling 19%. Varactor and mismatch losses were previously estimated and totaled 7-8%. The total estimated losses are approximately 35%.

The doubler was constructed with ℓ_1 having a Q of 115, ℓ_2 and ℓ_x having Q's of approximately 100. The circuit was tuned as follows. Eight milliwatts of power at 68 mcs. was applied to the doubler with the output fed to a spectrum analyzer (or receiver). Coil ℓ_2 was adjusted to null, insofar as possible, the 68 mcs. present in the output and henceforth is left as is until final trimming. The 136 mc output is then peaked by adjusting ℓ_x, ℓ_1 and the bias voltage. The adjustments of ℓ_x, ℓ_1 and E_b will have to be repeated several times.

In the above example the maximum output of 136 mcs ($2\omega_o$) was achieved with a bias voltage (E_b) of -5 volts and the slug in coil ℓ_x almost extracted from the coil. The efficiency was approximately 55%; the output power (136 mcs.) was 4.4 milliwatts and the 68 mcs (ω_o) feed thru was in excess of 30 db below the 136 mcs ($2\omega_o$). One turn was removed from coil ℓ_x and the doubler again tuned to achieve maximum output at 136 mcs ($2\omega_o$). The bias (E_b) was further reduced to -4.5 volts and the 136 mcs ($2\omega_o$) output was increased to 5 milliwatts, yielding an efficiency of approximately 62%. Spurious output frequencies included ω_o at 35 db down, $3\omega_o$ at 28 db down, $4\omega_o$ in excess of 50 db down, $5\omega_o$ in excess of 50 db down, and $6\omega_o$ in excess of 60 db down.

Using the same design procedure but with larger, higher Q coils (also toroids), doublers of approximately 85% efficiency have been constructed. However, their stability, rigidity and physical size were not suitable for this application. The use of slightly larger coil forms, perhaps 50% greater diameter, should probably yield an efficiency of approximately 70-75%; however, the available space did not permit the use of the larger size coils.



Varactor Impendance - Vs - Frequency
doubler
R out
Max. eff.
R in & R out
Max. P_o
R in
Max. eff.
10^3
100
10
1
xR_s
$10^{-4}W_c$
$10^{-3}W_c$
$10^{-2}W_c$
$0.1W_c$
W_c
Input Frequency W_o

Chart 1

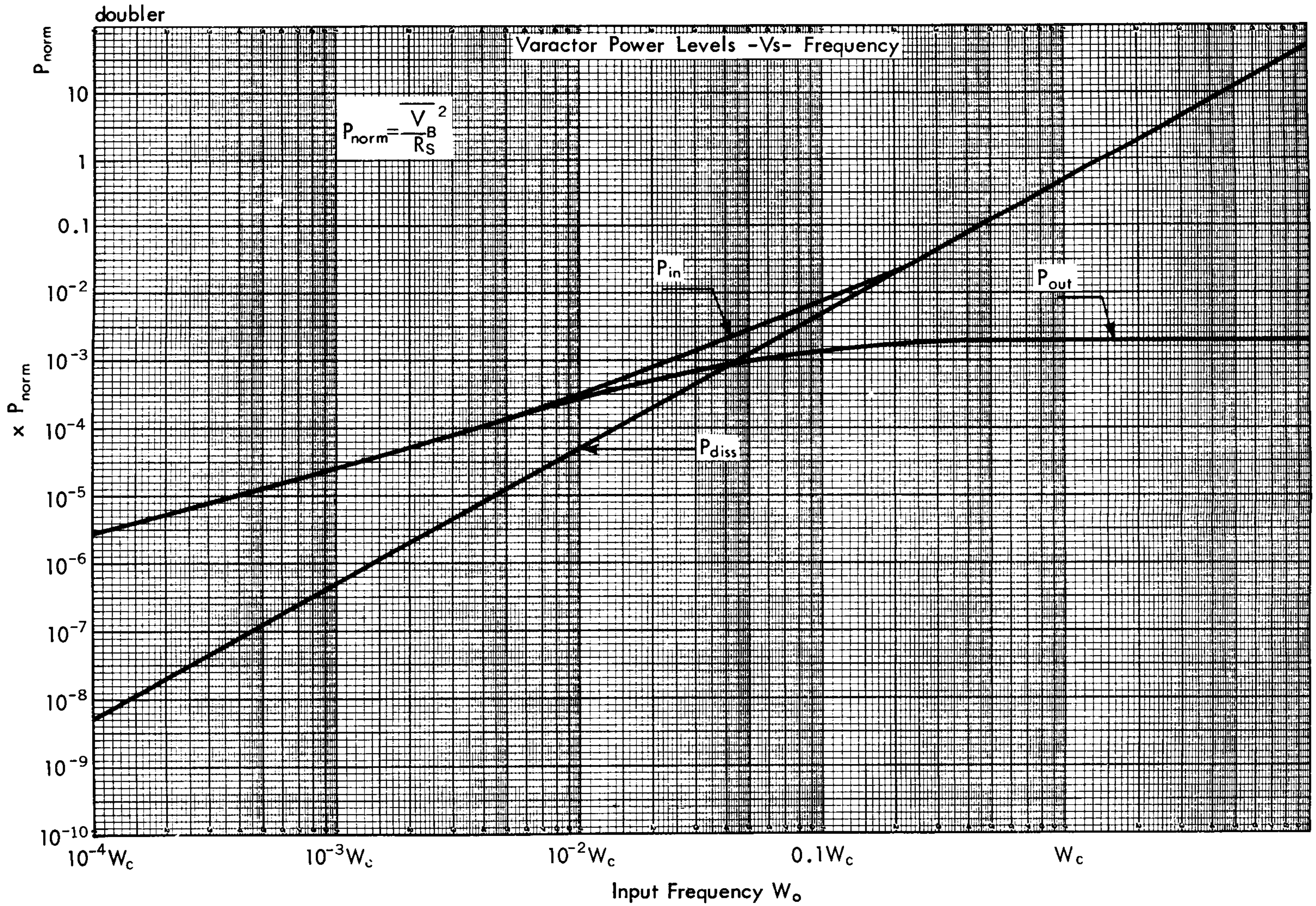

doubler
Varactor Power Levels -Vs- Frequency
$P_{norm} = \frac{\overline{V_B}^2}{R_S}$
P_{norm}
$\times P_{norm}$
10
1
0.1
10^{-2}
10^{-3}
10^{-4}
10^{-5}
10^{-6}
10^{-7}
10^{-8}
10^{-9}
10^{-10}
P_{in}
P_{out}
P_{diss}
$10^{-4}W_c$
$10^{-3}W_c$
$10^{-2}W_c$
$0.1W_c$
W_c
Input Frequency W_o

Chart 2

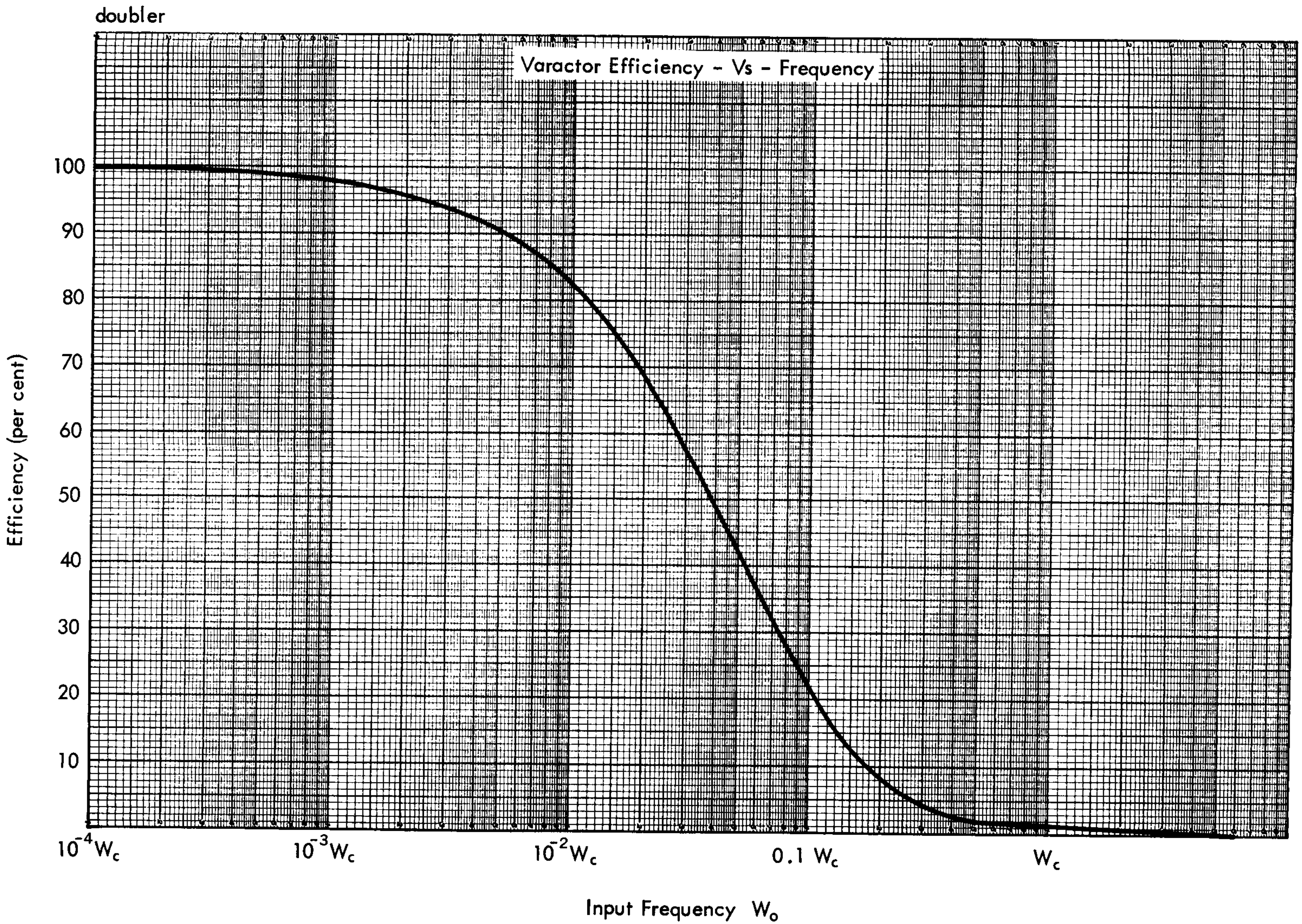
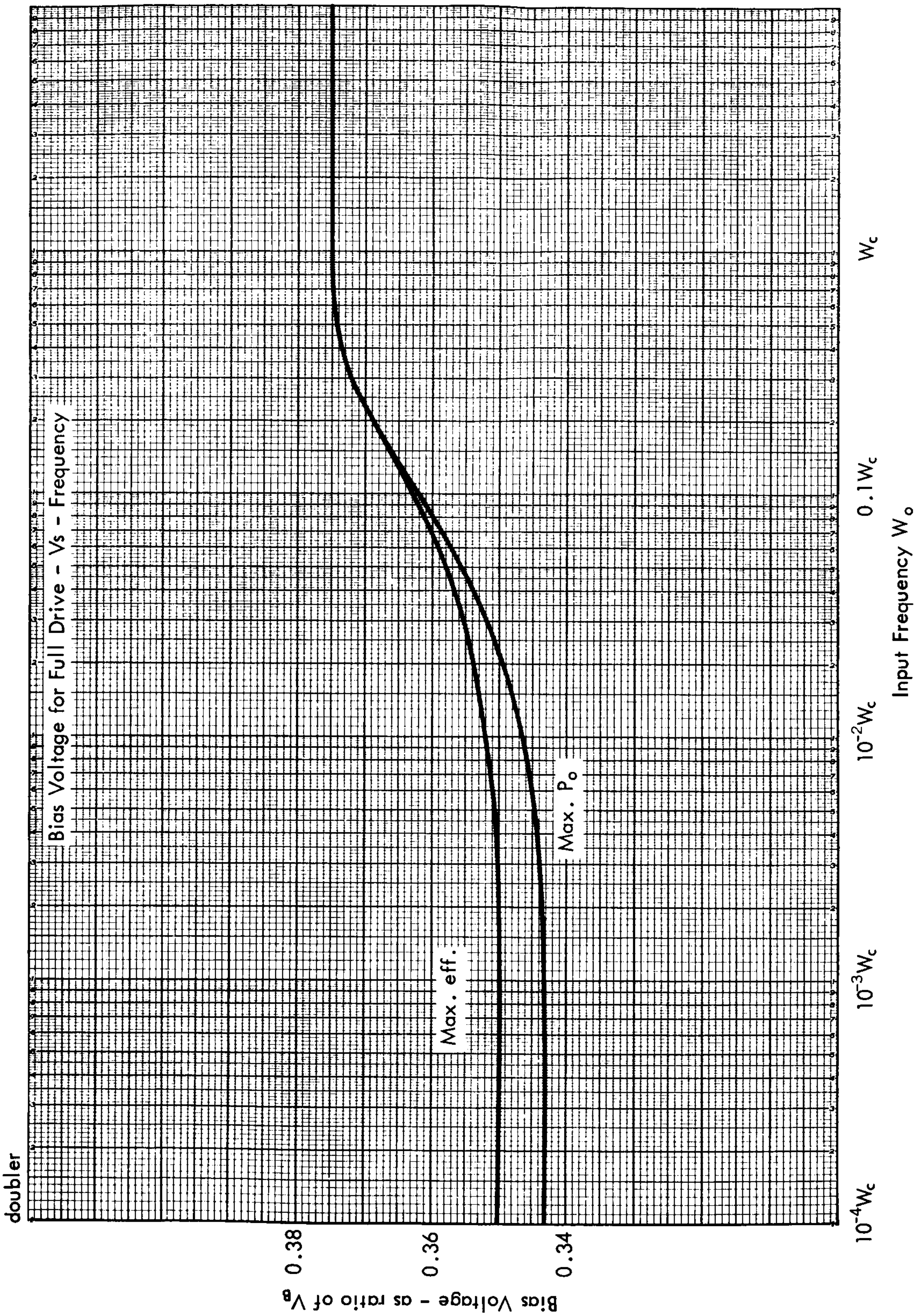

doubler
Varactor Efficiency - Vs - Frequency
100
90
80
70
60
50
40
30
20
10
Efficiency (per cent)
10^-4 W_c
10^-3 W_c
10^-2 W_c
0.1 W_c
W_c
Input Frequency W_o

Chart 3



Bias Voltage for Full Drive - Vs - Frequency
doubler
Max. eff.
Max. P_o
0.38
0.36
0.34
Bias Voltage - as ratio of V_B
$10^{-4}W_c$
$10^{-3}W_c$
$10^{-2}W_c$
$0.1W_c$
W_c
Input Frequency W_o

Chart 4

Design of a Single Ended Frequency Tripler

Specifications

$$\omega_o = 22.67 \text{ mcs.}$$

$$3\omega_o = 68 \text{ mcs.}$$

$$P_{in} = 12 \text{ milliwatts}$$

$$R_{in} = R_{out} = 50 \text{ ohms}$$

The design of a frequency tripler is similar to that for a frequency doubler. The varactor will provide a $2\omega_o$ voltage upon being driven with a ω_o current. If we permit a $2\omega_o$ current to flow (idler current) through a minimum impedance we will also have available the sum of ω_o and $2\omega_o$ — or a $3\omega_o$ voltage. ($4\omega_o$ is also available if we would provide a path for it.) Our problem then is to drive the varactor efficiently (low input reactance to ω_o); provide a low impedance path for a $2\omega_o$ current; and, provide a low reactance output path for $3\omega_o$. At the same time we desire to minimize ω_o and $2\omega_o$ currents in the output, and, to block $3\omega_o$ power from being fed to the ω_o source or to be shunted by the $2\omega_o$ (idler) circuit. These are general requirements for a frequency tripler and a suitable circuit can be arrived at by using techniques similar to those used to design the frequency doubler. However, we shall use the tripler to drive the doubler to achieve a ×6 multiplier circuit, and our requirements will be more specific. Referring to the doubler circuit just designed, the two traps, ω_o and $2\omega_o$, because of their relatively low values of reactance at lower frequencies, will provide very little isolation to our new ω_o of 22.67 mcs. and $2\omega_o$ of 45.33 mcs. Therefore, the output circuit of our tripler will be required to present a suitable attenuation at these frequencies—approximately 10 db at $2\omega_o$ (45 mcs.) and 20 db at ω_o (22.67 mcs.). A possible alternative could have been to convert either one or both of the traps in the previously designed doubler to series resonant circuits. However, as previously noted, spurious outputs would doubtless have been a problem for our specific purpose.

We shall use the same procedure to select a varactor for the tripler as for the doubler. Briefly, we desire approximately 50 ohms input and output impedances; and, we should drive the varactor for a large portion of its permissible voltage swing. Our input frequency will be 22.67 mcs. (ω_o), so our normalized input frequency will fall in the range of $2 \times 10^{-3}\,\omega_c$ to $3 \times 10^{-4}\,\omega_c$ if we assume we have varactors in a range of 10 gc to 70 gc cut-off frequency. We can make a simple table similar to Table 1. Table 2 shows the various parameters we shall look for in varactors of different cut-off frequencies (ω_c).

Table 2

($P_{in} \cong$ 12 milliwatts)

ω_c	70 gc	30 gc	10 gc
ω_o	$3 \times 10^{-4}\,\omega_c$	$7.5 \times 10^{-4}\,\omega_c$	$2 \times 10^{-3}\,\omega_c$
P_n	1700 watts	600 watts	240 watts
R_s	0.16 ohm	0.4 ohm	1.1 ohm

The P_{norm} is determined from Chart No. 5; the R_{in} and R_{out} are determined from Chart No. 6. Since the optimum doubler input impedance was 43 ohms, we can figure on aiming for a 43-ohm tripler output

(R_3). If we assume a low impedance for second harmonic current flow $(R_2 \leq 5R_s)$, we can estimate the varactor efficiency from Chart No. 7 to be between 88% and 98% depending on the quality (ω_c) of the varactor we use. It is important with respect to varactor effficiency to use low-loss idler paths. In higher order frequency multipliers the idler circuit, or circuits, are the most critical areas in regard to circuit band widths and conversion efficiencies.

A varactor was found in stock with the following characteristics:

$$f_c = 45 \times 10^9 \text{ cycles}$$

$$P_{norm} = 10^3 \text{ watts}$$

$$R_s = 0.32 \ \Omega$$

$$C_6 = 15.4 \ \mu\mu f$$

$$C_{18} = 11.5 \ \mu\mu f$$

$$V_B = 18 \text{ volts}$$

and:

$$\omega_o \text{ (normalized)} = \frac{f_{in}}{f_c} = \frac{22.67 \times 10^6}{45 \times 10^9} = \underline{5 \times 10^{-4} \ \omega_c}$$

$$P_{in} \cong 10^{-5} \ Pn \qquad \text{(from Chart No. 5)}$$

$$P_{in} = 10^{-5} \times 10^3 \cong \underline{10^{-2} \text{ watts}} - \text{(ok)}$$

$$R_{in} = 280 R_s = 280 \times 0.32 \cong \underline{90 \ \Omega} \qquad \text{(from Chart No. 6)}$$

$$R_{out} = 130 \ R_s = 130 \times 0.32 \cong \underline{42 \ \Omega}$$

The one parameter that is somewhat out of line is the input impedance of approximately 90 ohms which will cause some power loss and will also cause a voltage reflection to the source generator. A matching section could be used but would contribute an efficiency loss of several percent. We will try to achieve a degree of correction to the expected mismatch by altering the value of the input reactance after the tripler is constructed.

The following circuit shall be used to make our tripler compatible with the previously designed doubler.

Note: We should not necessarily conclude that a cascaded tripler-doubler is superior to a single-X6 multiplier. Also, as the operating frequency becomes very small relative to the varactor cut-off frequency (ω_c), the varactor losses become insignificant compared to circuit losses; and, providing full varactor drive is of course not as important to overall circuit efficiency as it may be at higher relative operating frequencies.

We shall make both shunt legs ($2\omega_o$ current path) of the below circuit resonant to $2\omega_o$ (45 mcs.). The $\ell_2 - c_2$ leg ($2\omega_o$ idler) will have higher reactance elements than the $\ell_x - c_x$ leg to reduce leakage losses at ω_o and $3\omega_o$. From an efficiency standpoint it may be desirable to tune the $\ell_x - c_x$ leg to a different frequency and then series resonanate the $2\omega_o$ current component by tuning $\ell_2 - c_2$ accordingly; however, ultimately it is our purpose to use the resulting X6 output as a local oscillator and it is very important that the signal be "clean." The condition under

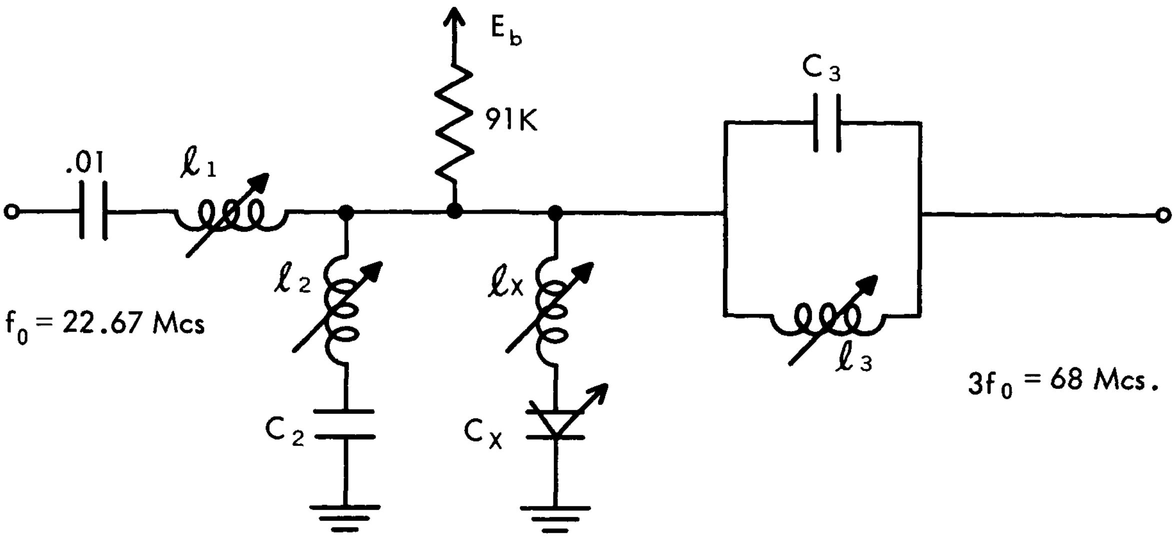


Figure 3—Tripler Circuit

which both shunt legs are series resonant to $2\omega_o$ results in a minimum of $2\omega_o$ power in the output. In the event a significant $2\omega_o$ signal feeds into the subsequent doubler, a large number of spurious signals will be present in the final output due to mixing action in the doubler. It was necessary in this case that the two shunt legs were carefully tuned to minimize $2\omega_o$ in the tripler output.

We shall now determine the component values necessary to achieve optimum circuit operation in conjunction with the selected varactor.

$\ell_x - c_x$ to be series resonant at 45.34 mcs. c_x is to be biased at approximately $V_B/3$ (Chart No. 8) or 6 volts

$$\underline{C_{x(6)} = 15\ \mu\mu f}$$

$$\underline{\therefore \ell_x = 0.8\ \mu h}$$

$$\left(\omega_{(45)} = \frac{1}{\sqrt{\ell_x c_x}}\right)$$

The $\ell_x - c_x$ reactance at

$$\omega_o \cong +j115 - j460 \cong \underline{-j345}$$

$$3\omega_o \cong +j350 - j150 \cong \underline{+j200}$$

The input coil (ℓ_1) is now selected to resonate with $\ell_x - c_x$ at ω_o by making ℓ_1 equal to $+j$ 345 ohms at ω_o. Bear in mind that ℓ_1 will later be altered to improve the input impedance match. Similarily, the output trap $\ell_3 - c_3$ which is parallel resonant to ω_o must present a series reactance of $-j$ 200 ohms at $3\omega_o$ to provide, in conjunction with $\ell_x - c_x$, a low reactance path for the output power.

$$\omega_o \ell_1 = +j345 \text{ ohms}$$

$$\underline{\ell_1 = 2.4\ \mu h}$$

also

$$3\omega_o \ell_1 \cong +j1000 \text{ ohms}$$

To determine ℓ_3 and c_3 set up two simple equations:

1) $+j\omega_o \ell_3 - j\dfrac{1}{\omega_o c_3} = 0$ (trap for ω_o)

and

2) $$\frac{(+j3\omega_o \ell_3)\left(-j\dfrac{1}{3\omega_o c_3}\right)}{+j3\omega_o \ell_3 - j\dfrac{1}{3\omega_o c_3}} = -j200$$

(to series resonate with $\ell_x - c_x$ at $3\omega_o$)

solve the first equation:

$$\omega_o \ell_3 = \frac{1}{\omega_o c_3}$$

and substitute into the second:

$$\underline{c_3 \cong 12.5\ \mu\mu f}$$

$$\underline{\ell_3 \cong 3.8\ \mu h}$$

(See appendix for similar example)

Select the values of ℓ_2 and c_2 to be series resonant at $2\omega_o$ (approximately 45 mcs.) and to have larger reactance values than ℓ_x and c_x.

let $\underline{\ell_2 = 2.0\ \mu h}$

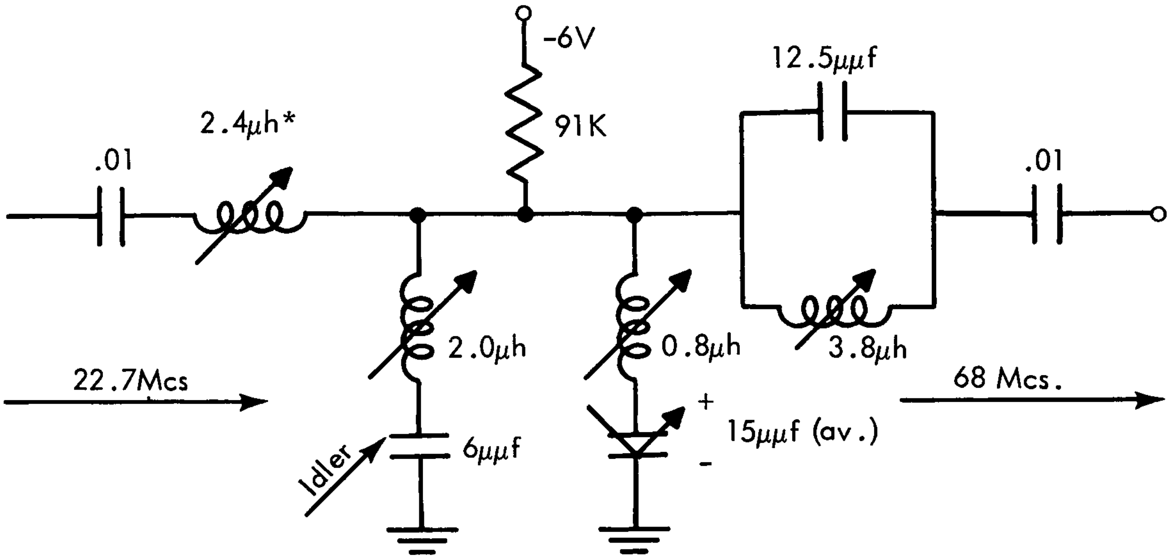
then: $\underline{c_2 = 6.0\ \mu\mu f}$

A check of the shunting effect of $\ell_2 - c_2$:

$$\omega_o)\ +j300 - j1150 \cong -j850 \text{ ohms}$$

$$3\omega_o)\ +j900 - j380 \cong +j520 \text{ ohms}$$

Note: At ω_o the $-j850$ ohm shunt is in series with ℓ_1 reactance of $+j$ 345 ohms. At $3\omega_o$ the $+j520$ ohms is in series with $\ell_x - c_x$ reactance of $+j200$ ohms. The shunting losses due to the idler leg should be in the neighborhood of 2%.

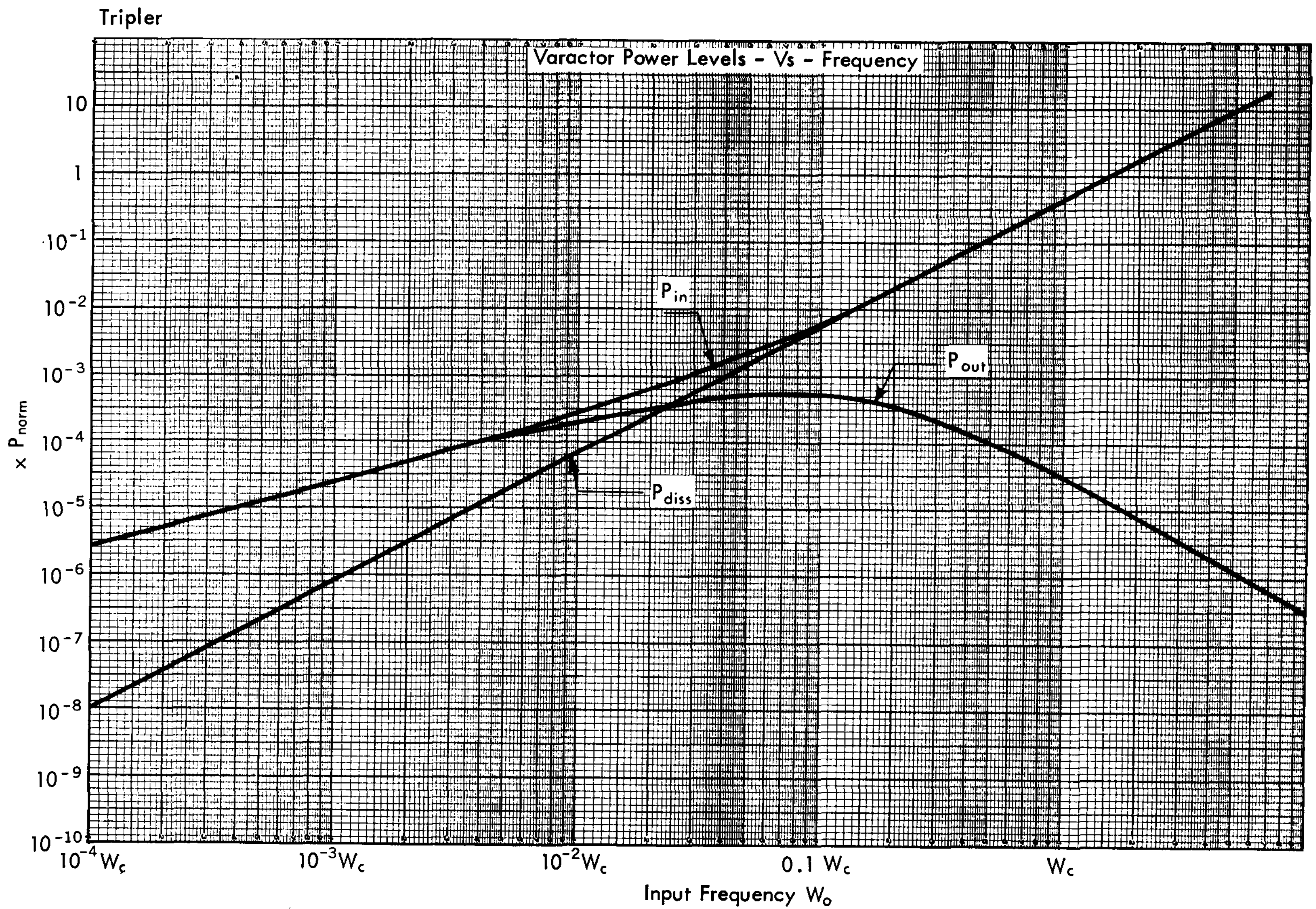


* To be altered during tuning for impedance matching

Figure 4–Component Values–Tripler

Times Six Cascade

The 22.67 mcs. to 68 mcs. tripler was coupled to the previously designed doubler for tuning. First, with bias applied, coils ℓ_x, ℓ_1, ℓ_2 and ℓ_3 are adjusted to produce a 136 mcs. signal at the output of the doubler. At this point the 136±45 mcs. will be rather large and ℓ_2 and ℓ_x are adjusted to minimize these signals as ℓ_1 and ℓ_3 are adjusted to enhance 136 mcs. Coil ℓ_3 is also adjusted to minimize 136±22 and 22 mcs., which are normally rather low due to the doubler reactances. The 136 mcs. output is further enhanced by readjustment of the coils and the bias voltage. At this point the 136 mcs. output was down approximately 14 db, due largely to the input impedance mismatch.



Tripler
Varactor Power Levels - Vs - Frequency
10
1
10^-1
10^-2
10^-3
10^-4
10^-5
10^-6
10^-7
10^-8
10^-9
10^-10
x P_norm
P_in
P_out
P_diss
10^-4 W_c
10^-3 W_c
10^-2 W_c
0.1 W_c
W_c
Input Frequency W_o

Chart 5

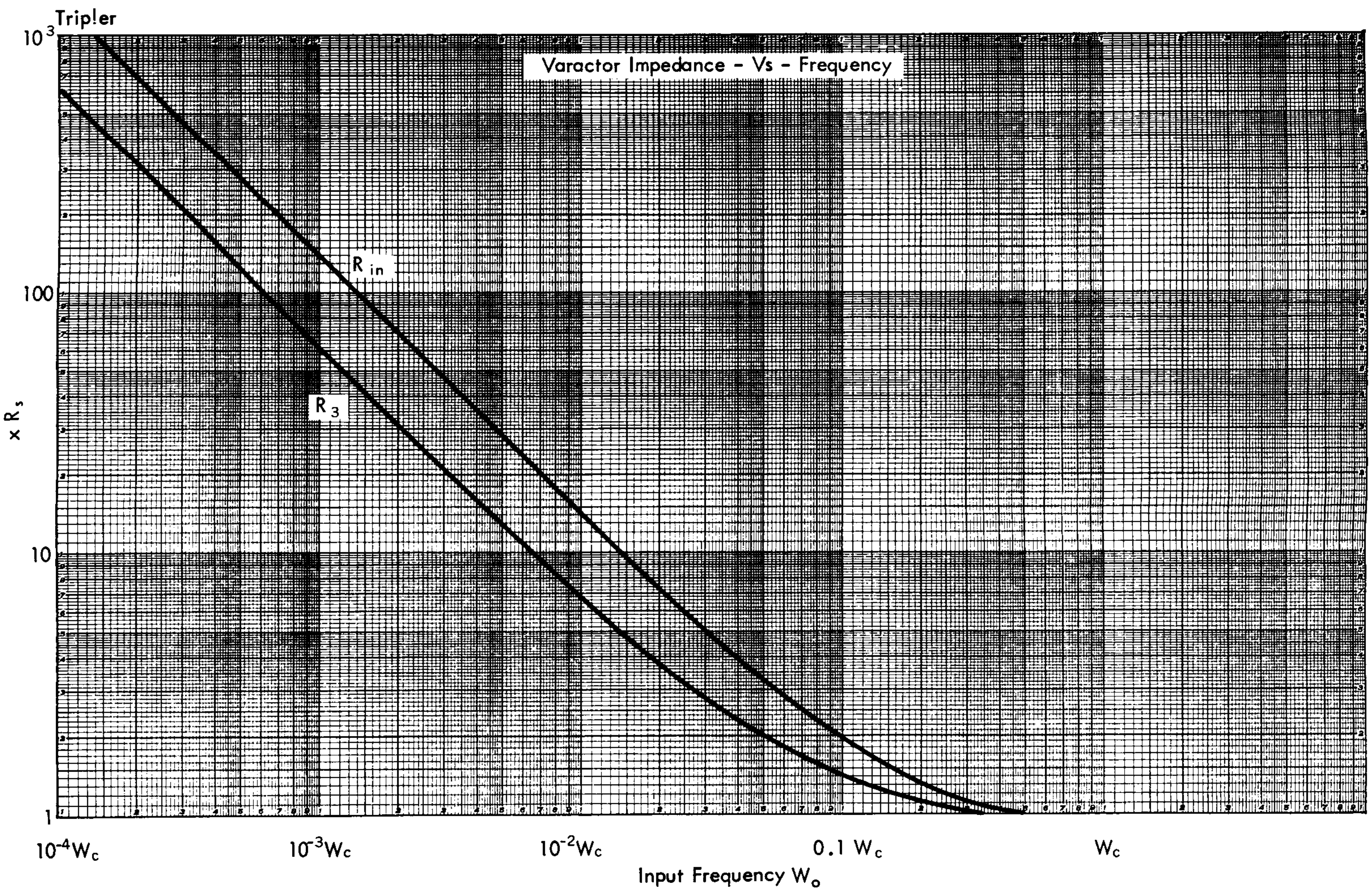

Tripler
Varactor Impedance - Vs - Frequency
R_{in}
R_3
10^3
100
10
1
$\times R_s$
$10^{-4}W_c$
$10^{-3}W_c$
$10^{-2}W_c$
$0.1\ W_c$
W_c
Input Frequency W_o

Chart 6

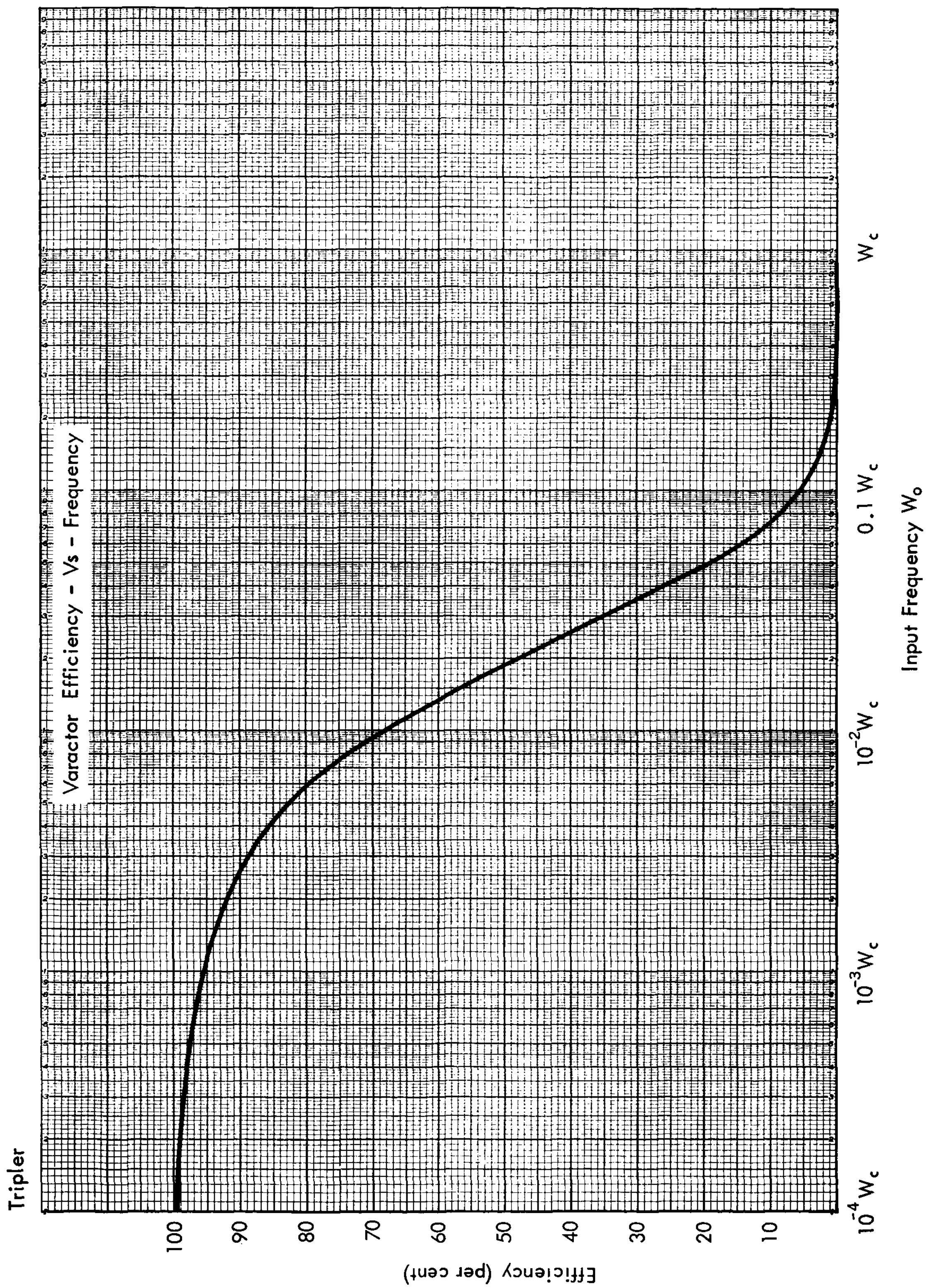

Tripler
Varactor Efficiency - Vs - Frequency
Efficiency (per cent)
100
90
80
70
60
50
40
30
20
10
10^{-4} W_c
10^{-3} W_c
10^{-2} W_c
0.1 W_c
W_c
Input Frequency W_o

Chart 7

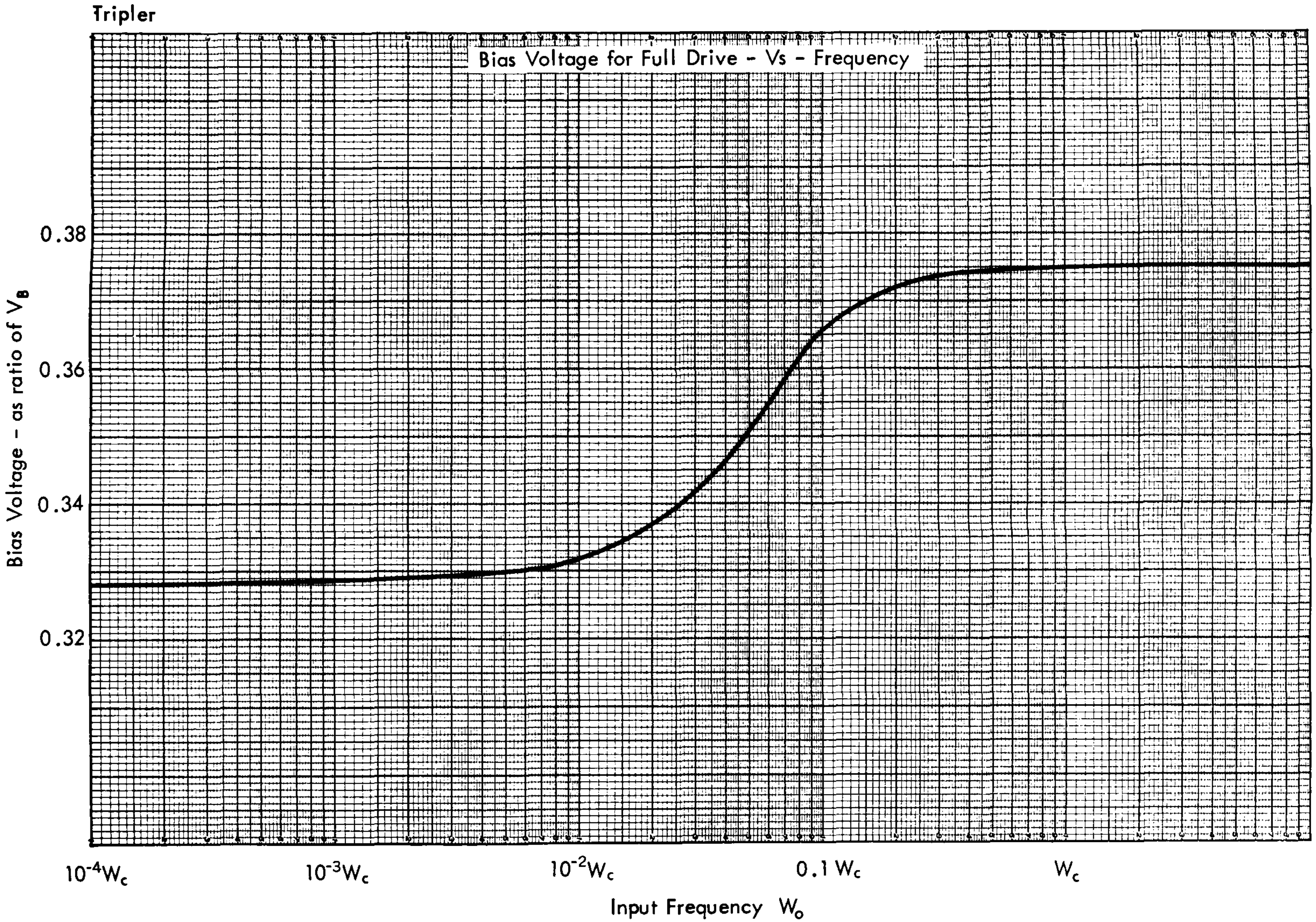

Tripler
Bias Voltage for Full Drive - Vs - Frequency
0.38
0.36
0.34
0.32
Bias Voltage - as ratio of V_B
$10^{-4}W_c$
$10^{-3}W_c$
$10^{-2}W_c$
$0.1\,W_c$
W_c
Input Frequency W_o

Chart 8

Note: All the excess loss (12 db) over the doubler loss of approximately 2 db was not due entirely to the tripler. As the input power to the doubler is reduced significantly from the design value its loss increases and perhaps the doubler loss in this instance is 5 or 6 db rather than the previously measured 2 db.

At this point several turns were removed from coil ℓ_1 as the core had been removed for maximum 136 mcs. output. After tuning, the core was still removed for maximum output and two additional turns were removed. Tuning was again accomplished and an output of 136 mcs. approximately 5 db down in power from the 22 mcs. input power was achieved. The resulting inductance of coil ℓ_1 was not measured but was estimated to be approximately 1-1/2 μh. The bias voltage on the tripler stage was -5 volts, the same as for the doubler, and only slight adjustment was made to the doubler circuitry of the multiplier.

With an input of 22.67 mcs. at 12 mw to the times six multiplier, the output consisted of 136 mcs. at 3.8 mw (approximately 5 db down) and a number of spurious responses. The output is shown in Table 3.

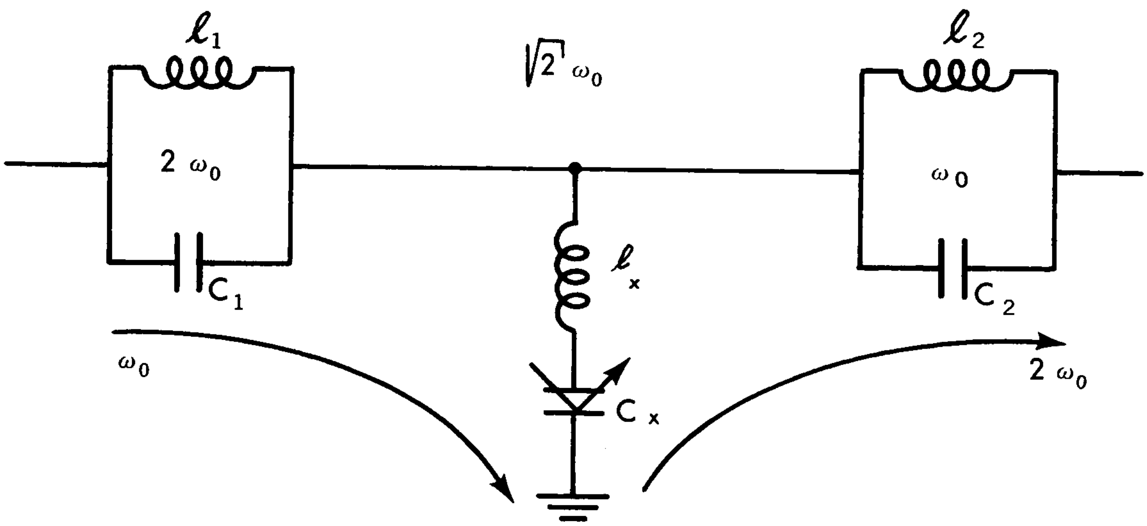
The requirement which led to the design of the ×6 multiplier was the need of a 136 mcs. local oscillator signal which was coherent with a transmitter frequency. In this case they were both derived from a 22.67 mcs. xtal oscillator. The unit described was assembled on a printed circuit board and high efficiency was not an over riding consideration. The package was somewhat limited in size and therefore rather small coils were utilized. Also it was important that spurious outputs be down significiantly. In conjunction with the same project a × 12 multiplier was bread boarded for an input frequency of 11.34 mc. and output of 136 mcs. This unit used larger size coils and consisted of two doublers and one tripler cascaded and constructed similarily to the X6 multiplier. The 136 mcs. output in the X12 multiplier was down approximately 6 db as compared to the 5 db in the X6 multiplier.

Table 3

Output Spectrum
(Power levels are relative to 136 mcs.)

f (mcs.)	Power Out
f_o 22.7	-33db
$2 f_o$ 45	-35db
$3 f_o$ 68	-32db
$4 f_o$ 91	$>$ -50db
$5 f_o$ 113	-36db
$6 f_o$ 136	0db (reference)
$7 f_o$ 158	-34db
$8 f_o$ 181	-42db
$9 f_o$ 204	-28db
$10 f_o$ 227	$>$ -60db
$11 f_o$ 249	$>$ -60db
$12 f_o$ 270	$>$ -50db

Appendix



Determine ℓ_1 and ℓ_2 in terms of ℓ_x under the following conditions:

The input trap ($\ell_1 - c_1$) is paralled resonant at $2\omega_o$

The output trap ($\ell_2 - c_2$) is parallel resonant at ω_o

The input ($\ell_1 - c_1$ and $\ell_x - c_x$) is series resonant at ω_o

The output ($\ell_x - c_x$ and $\ell_2 - c_2$) is series resonant at $2\omega_o$

The $\ell_x - c_x$ leg is series resonant at $\sqrt{2}\omega_0$

(1) (parallel resonant input trap)

$$+ j\, 2\omega_o \ell_1 - j \frac{1}{2\omega_o c_1} = 0$$

$$\underline{\frac{1}{\omega_o c_1} = 4\,\omega_o \ell_1}$$

(2) (series resonant input– $\ell_1 - c_1$ and $\ell_x - c_x$)

$$\frac{(+j\,\omega_o \ell_1)\left(-j\,\dfrac{1}{\omega_o c_1}\right)}{+j\,\omega_o \ell_1 - j\,\dfrac{1}{\omega_o c_1}} + j\,\omega_o \ell_x - j\,\frac{1}{\omega_o c_x} = 0$$

now set ℓ_x and c_x series resonant at $\sqrt{2}\,\omega_o$ and solve for $\omega_o C_x$

(3) $+j\sqrt{2}\,\omega_o \ell_x - j\dfrac{1}{\sqrt{2}\,\omega_o c_x} = 0$

$$\underline{\frac{1}{\omega_o c_x} = 2\,\omega_o \ell_x}$$

now substitute (1) and (3) into (2)

$$\frac{(+j\,\omega_o \ell_1)(-j\,4\omega_o \ell_1)}{+j\,\omega_o \ell_1 - j\,4\omega_o \ell_1} + j\,\omega_o \ell_x - j\,2\omega_o \ell_x = 0$$

$$\frac{(+j\,\omega_o \ell_1)(-j\,4\omega_o \ell_1)}{-j\,3\omega_o \ell_1} - j\,\omega_o \ell_x = 0$$

$$+\frac{j\,4\omega_o \ell_1}{3} - j\,\omega_o \ell_x = 0$$

$$\underline{\ell_1 = 3/4\,\ell_x}$$

Similiarily for ℓ_2

(4) series resonant output ($\ell_x - c_x$ and $\ell_2 - c_2$)

$$+j\,2\omega_o \ell_x - j\,\frac{1}{2\omega_o c_x} + \frac{(+j\,2\omega_o \ell_2)\left(-j\,\dfrac{1}{2\omega_o c_2}\right)}{+j\,2\omega_o \ell_2 - j\,\dfrac{1}{2\omega_o c_2}} = 0$$

(5) parallel resonant output trap ($\ell_2 - c_2$)

$$+ j\, \omega_o \ell_2 - j \frac{1}{\omega_o c_2} = 0$$

$$\underline{\frac{1}{\omega_o c_2} = \omega_o \ell_2}$$

now substitute (3) and (5) into (4)

$$+ j\, 2\omega_o \ell_x - j\, \omega_o \ell_x + \frac{(+ j\, 2\omega_o \ell_2)\left(- j \frac{1}{2} \omega_o \ell_2\right)}{+ j\, 2\omega_o \ell_2 - j \frac{1}{2} \omega_o \ell_2} = 0$$

$$+ j\, \omega_o \ell_x + \frac{(+ j\, 2\omega_o \ell_2)\left(- j \frac{1}{2} \omega_o \ell_2\right)}{+ j \frac{4\omega_o \ell_2 - \omega_o \ell_2}{2}} = 0$$

$$+ j\, \omega_o \ell_x + \frac{(+ j\, 2\omega_o \ell_2)\left(- j \frac{1}{2} \omega_o \ell_2\right)}{+ j \frac{3}{2} \omega_o \ell_2} = 0$$

$$+ j\, \omega_o \ell_x + \frac{4}{3}\left(- j \frac{1}{2} \omega_o \ell_2\right) = 0$$

$$+ j\, \omega_o \ell_x - j \frac{2}{3} \omega_o \ell_2 = 0$$

$$\underline{\ell_2 = \frac{3}{2} \ell_x}$$

For C_1:

from (1)

$$\frac{1}{\omega_o c_1} = 4\omega_o \ell_1$$

$$\frac{1}{\omega_o c_1} = 3\omega_o \ell_x \quad \left(\ell_1 = \frac{3}{4} \ell_x\right)$$

$$\frac{1}{\omega_o c_1} = \frac{3}{2\omega_o c_x} \quad \text{(from 3)}$$

$$\underline{c_1 = \frac{2}{3} c_x}$$

For c_2

from (5)

$$\frac{1}{\omega_o c_2} = \omega_o \ell_2$$

$$\frac{1}{\omega_o c_2} = \frac{3}{2} \omega_o \ell_x \quad \left(\ell_2 = \frac{3}{2} \ell_x \right)$$

$$\frac{1}{\omega_o c_2} = \frac{3}{4 \omega_o c_x} \quad \text{(from (3))}$$

$$\underline{c_2 = \frac{4}{3} c_x}$$